Method for dividing the bit rate of QPSK signals into two or more sub-channels

The invention relates to a method for dividing the bit rate of QPSK signals into at least two sub-channels having band width limited filters in the modulator and the demodulator, by means of splitting the spectrum of the QPSK signals.

The basics of the PSK method are known from the technical manual "Nachrichtentechnik" [Communications technology] by E. Herter/W. Lörcher, 5th edition, which appeared in the Hanser Verlag [publishing company] in 1990, pages 110 ff., and the implementation of PSK modulators and demodulators and frequency multiplication were described there. Thus it is possible to generate a carrier $2f_T$ from a 2-PSK signal, by means of squaring, from which the desired carrier f_T results afterwards, by means of frequency division. For this, it is indicated that in general, squaring has to occur in the case of an N-PSK signal *ld(n)md*. During squaring, the phase angles are doubled. After the first squaring step at 2-PSK, the signal therefore is given the phase position 0 and 360°. But since these phase positions are the same, the spectrum of the signal that has been squared twice

contains contributions, after the phase angle doubling, which point in the same direction. Seen spectrally, this means that the desired line is reached at a multiple of the original carrier frequency f_T, for example at four f_T. The reference carrier of the frequency f_T that is obtained by means of frequency division, in this connection, has a phase that is displaced by $n \times \frac{\pi}{2}$ $(n = 0 \ldots 3)$, as compared with the correct zero phase.

From the IEEE Transactions on Communications 37, No. 5 (May 1989), pages 437 to 448, a proposal is known how the bit rate of QSPK can be doubled by adding a second orthogonal signal. Fig. 4 on page 447 shows such signal shapes. Because of the perpendicular flanks of the pulses, the band width is very great, i.e. the orthogonality is lost when the band width is limited, and inter-symbol interference (ISI) and cross-talk (ÜS) occur between the channels. At the end of the essay, the authors, D. Saha and G. Birdsall, discuss systems limited in band width, which use band width limited transmission filters $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ and corresponding matched filters $\boldsymbol{P}_1$* and $\boldsymbol{P}_2$* on the receiving side (Fig. 13 on page 446). The bit rate 1/T = $2f_g$ for one branch of a QPSK system (in other words a total of $4f_g$) is split into 1/2T twice there, and is therefore the same as for QPSK. This

arrangement is used for the sine carrier and the cosine carrier, in each instance. The authors make the statement that there are infinitely many possibilities for the pairs $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$, and give three examples in Figure 14 on page 447, without the related pulse responses of the individual filters $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$, the transmission and reception filters $\boldsymbol{P}_1\boldsymbol{P}_1^*$ and $\boldsymbol{P}_2\boldsymbol{P}_2^*$, and do not discuss the cross-talk $\boldsymbol{P}_1\boldsymbol{P}_2^*$. Since the filters $\boldsymbol{P}_1$ are real and $\boldsymbol{P}_2$ are imaginary, it holds true that $\boldsymbol{P}_1^* = \boldsymbol{P}_1$ and $\boldsymbol{P}_2^* = -\boldsymbol{P}_2$. A closer examination shows that the conditions free of ISI and ÜS can only be achieved with the examples (a) and (b), and that the example (c) according to Figure 14 does not fulfill the conditions, in disadvantageous manner.

The idea of adding a second pulse, orthogonal to the control pulse of the QPSK, for modulating the sine carrier and the cosine carrier, is also known from U.S. 4,680,777.

Proceeding from the state of the art according to the IEEE reference, the invention is based on the task of indicating a method that fulfills the conditions: free of inter-symbol interfrequency (ISI) and cross-talk (ÜS) between the channels, and reduces the infinitely many possibilities, as mentioned, to a class of filters, in its implementation.

The stated task is accomplished by the method indicated in claim 1, in the interaction of the individual method steps, and indicates the division of the spectrum of the QSPK signals into at least two frequency bands, the transmission of the same in frequency multiplex, and the dimensioning of the filters in the modulator and the demodulator as a function of the transmission function.

Advantageous further method steps and their embodiments are indicated, as a supplement, in the dependent claims.

The invention will be explained in detail in the following, based on the Figures 1 to 17 shown in the drawings.

In Figure 1, the orthogonal pulse shapes known from the IEEE reference mentioned initially are shown.

In Figure 2, the basic band model of a Q^2PSK transmitter and receiver, as it can be derived from Figure 13 of the IEEE reference, is shown using the example of a transmission and reception branch for an orthogonal carrier. On the transmitter side, a serial parallel conversion is performed first, and the converted signal is passed to the two filters P_1 and P_2. The

signal that has been split in this manner is passed to an addition stage after filtering, modulated with a cosine carrier and, in a second, identical branch, with a sine carrier, and transmitted to the receiver with cosine and sine demodulators. The demodulated signals go to the two signal branches having filters $\boldsymbol{P}_1$* and $\boldsymbol{P}_2$*, are scanned with multiples of f_T• and decided for the data signals in a threshold decision.

In Figure 3, the examples that are evident from the IEEE reference, Figure 14, mentioned initially, are reproduced. Specifically, in Figures 3a, 3b, and 3c, the pulse responses to the examples of filter pairs a, b, c in Figure 3 are shown, namely divided according to transmission filter and pulse response of the overall system, and the cross-talk behavior. Figure 3c shows that the requirements: free of inter-symbol interfrequency (ISI) and cross-talk (ÜS) are not precisely fulfilled.

The consideration of the invention using Figures 4 to 17 proceeds from these known systems.

The filters used for signal splitting can be filters that are adjacent to one another in terms of frequency (Variant A), or

filters that lie in the same frequency range (Variant B). The method according to the invention furthermore solves the use of a duobinary coding. Furthermore, the method according to the invention can be expanded from a draft method of two (***Q^2PSK***) to ***n*** (***Q^nPSK***) partial signals.

Furthermore, the open question of the use of a duobinary coding is solved using the method. By replacing the filters $\boldsymbol{P}_2 \ldots \boldsymbol{P}_n$ with a serial circuit of a low-pass filter $\boldsymbol{P}_1$ and subsequent modulation with equidistant sine and cosine carriers, a multi-carrier system is obtained. Its implementation can take place similar to the case of OFDM (Orthogonal Frequency Division Multiplex) by way of DFT (Discrete Fourier Transformation) and IDFT (Inverse DFT). As compared with OFDM, however, ***Q^nPSK*** offers several advantages, namely a more compact spectrum, a lower crest factor, less sensitivity in the case of frequency-selective channels and with regard to carrier synchronization.

In the following, the design of Q^2PSK systems according to the invention will first be described.

We proceed from a QPSK having an ideal low-pass channel $\boldsymbol{H}_i$ having the band width ω_g according to Fig. 4, in which the signal

progressions of the filters and the individual design steps are indicated. The low-pass channel $\boldsymbol{H}_i$ can, as indicated with a broken line, be changed by means of a Nyquist flank at ω_g, for a practical implementation, without any change occurring at the zero passages of the pulse response at multiples of $1/2f_g$, as is evident from the diagram in the first line at the top and the pulse response next to it. Since transmission occurs at half the bit rate $1/f_g$ in the $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ branch (Fig. 2 or Fig. 7 or Fig. 8), a low-pass having half the band width can be used for $\boldsymbol{P}_1^2$ (second line in Fig. 4). In Fig. 4, $\boldsymbol{P}_1^2$ was already supplemented with a Nyquist flank at $\omega_g/2$. The zero places of the related pulse response accordingly lie at multiples of $1/f_g$ (representation line 1 in Fig. 4, on the right). If one forms

$$P_2P_2^* = H_i - P_1^2$$

the zero places of the pulse response belonging to $\boldsymbol{P}_{2m} = \boldsymbol{P}_2\boldsymbol{P}_2{*}$ also lie at $1/f_g$, so that transmission can also take place at the bit rate $1/f_g$ via this channel. It is evident from the second step in Fig. 4 that $\boldsymbol{P}_2\boldsymbol{P}_2{*}$ has the same Nyquist flank at $\omega_{g/2}$ as $\boldsymbol{P}_1^2$, so that in this range, $\boldsymbol{P}_1^2 + \boldsymbol{P}_2\boldsymbol{P}_2{*} = 1$.

The division of the PSK signal into the real P_1 and the purely imaginary P_2 is easily possible by means of adding the root sign $\sqrt{N}$ and the corresponding sign at P_2, thereby finding the desired pulse former pairs P_1 and P_2. When speaking about the pulses P_1 and P_2, these are the pulses that can be captured at the outputs of the filters P_1 and P_2, and the same also applies to P_1* and P_2*. If the method according to the invention is used, then no cross-talk caused by P_1P_2* or P_2P_1* occurs, either. This is discussed and documented in the lower part of Fig. 4. Because of the same Nyquist flanks at $\omega_g/2$, a cross-talk spectrum results that is symmetrical to $\omega_g/2$ and point-symmetrical to $\omega = 0$. Since P_1 and P_2 have orthogonal carriers (are in square with one another), this cross-talk can also be referred to as square cross-talk. This spectrum includes a pulse response that has zero places at the multiple of $1/f_g$ and therefore does not interfere in the scanning points of the wanted signal. This can be documented in that P_1P_2* can be imagined by conversion of a real spectrum $R(\omega)$ with sin $\frac{\omega_g}{2}t$.

By means of this sine carrier, zero places then occur in the related time signal, at the multiple of $1/f_g$. This adjacent arrangement of the filters is referred to as Variant A.

Furthermore, a special case is shown in Fig. 4, at the bottom. Specifically, if the Nyquist flank runs vertically, the two transmission channels are separated by frequency multiplex, as is evident from the last diagram. However, for an implementation, the vertical flank of $\boldsymbol{P}_2$ at the border frequency ω_g is disruptive. Proceeding from a Nyquist flank of $\mathbf{H}_i$, it is possible, as is evident from Fig. 5, in which the conditions for avoiding cross-talk (ÜS) are indicated, to apply a Nyquist flank also at ω_g in the case of $\boldsymbol{P}_{2m} = \boldsymbol{P}_2\boldsymbol{P}_2{}^*$. This reaches down into the region below ω_g. In order to avoid cross-talk (ÜS), in this case $\boldsymbol{P}_1$ is not allowed to drop down into the region of this Nyquist flank.

In this way, it is assured that the two channels are separated by means of frequency multiplex, but are allowed to overlap with their Nyquist flanks at $\omega_g/2$, without any cross-talk (ÜS) occurring.

Fig. 5 indicates the conditions for avoiding cross-talk (ÜS) between the PSK signals. It can be shown that Fig. 3a can also be interpreted as a special case of the design method, if the additive Nyquist flanks $\boldsymbol{P}_a$ are selected accordingly.

However, the question comes up whether the known example in Fig. 3a is particularly advantageous, since the two channels utilize the entire band width. In the case of a power-limited transmission channel, the interference intervals $\frac{E_b}{N_o}$ are equal, in the example according to Fig. 3a and in the design method corresponding to the filter pair example in Fig. 3b (and, of course, in the case of all other channels having Nyquist flanks at $\omega_g/2$). The filter pairs $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ are, as is also indicated in the state of the art in the IEEE, multiplied with the factor $\sqrt{2}$ during division into the lower and higher frequency range, with overlapping Nyquist flanks at $\omega_g/2$, in order to make $\frac{E_b}{N_o}$ the same as in QPSK. The peak amplitude is then smaller as compared with the example of Fig. 3a, which brings about a gain in the case of amplitude-limited channels, thereby documenting that the example of Fig. 3a is not advantageous. However, a gain occurs in the case of an expansion to multi-carrier systems of the Variant A. In the case of a roll-off factor of r=o, this is 3 dB. The example in Fig. 3a, with the greater peak amplitude, corresponds to a Variant B in the case of multi-carrier systems. For an implementation, the example of Fig. 3a would have to be freed of the perpendicular flanks. This is not possible without the occurrence of ISI and/or cross-talk.

Fig. 6 shows possibilities of a method implementation of the filtering of the signals $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ without cross-talk, and the transition to a multi-carrier system (Variant B).

In Example d, $\boldsymbol{P}_1$ is given a root Nyquist flank at ω_g, and $\boldsymbol{P}_2$ is given root flanks at $\frac{1}{4}\omega_g$ and $\frac{3}{4}\omega_g$. As a result, the cross-talk is zero, because the spectrum of $\boldsymbol{P}_1\boldsymbol{P}_2^*$ is symmetrical to $\omega_g/2$ and point-symmetrical to $\omega=0$ (see also Fig. 4, bottom). In this connection, it is practical if $\boldsymbol{P}_1^2$ and $\boldsymbol{P}_{2m}$ are converted to the base band by means of demodulation. For avoiding cross-talk, it is important that $\boldsymbol{P}_2$ is symmetrical around $\omega_g/2$ in the region of $\boldsymbol{P}_1$.

Example e shows that $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ can also be made equal in amount in this region. Additional channels having the same flanks can be added in the frequency multiplex. In this manner, one arrives at a multi-carrier system. In order for no cross-talk to occur, the individual channels must be separated in terms of frequency, in other words they are not allowed to overlap, at first, as is shown in Example f.

The root Nyquist flanks can also overlap, in terms of frequency, as shown in Example g. However, in this case, not only a square cross-talk but also an in-phase cross-talk occurs, which can be made to zero by means of an offset of adjacent channels by half a bit duration in the scanning time point (OQPSK).

This filter arrangement in the same frequency range will be called Variant B. As compared to Variant A, there is no advantage in the total bit rate, as was explained in connection with the explanation of Fig. 3a, which corresponds to Variant B. The filters in each channel (real or imaginary) form a Hilbert pair, as is known from the IEEE article. In the case of an implementation by means of modulation, it is recommended to convert with a carrier in the band center (two-sided band transmission). Variant B is known as a multi-carrier system from [4] and [5].

Expansion to duobinary transmission:

The expansion to partial-response or duobinary transmission is quite simple, according to the invention, taking into consideration the formation of a partial-response signal. It is known that the cosine crest channel $H_c(\omega)$ indicated in Fig. 7, top, supplies a corresponding pulse response. Transmission

occurs by way of the cosine crest channel at the bit rate $2f_g$, as in the case of the ideal low-pass. Its pulse response can, as shown in Fig. 7, be indicated as two pulse responses of an ideal low-pass, multiplied by the factor ½, which are offset relative to one another by the time $1/2f_g$, in other words the interval between the zero places of the sine function. With this, this pulse response that belongs to $\boldsymbol{H}_c(\omega)$ again has zero places at the interval of $1/2f_g$, as is evident from the diagrams below the block schematic. Practically, two Dirac surges that follow one another at an interval of $1/2f_g$ are transmitted, instead of one Dirac surge $\delta(t)$. Now, scanning again takes place at the interval of $1/2f_g$ at the receiver, but offset by $1/4f_g$ as compared with the ideal low-pass. As a result, one obtains the values ½, according to Fig. 7, at $\pm 1/4f_g$ of the pulse response $\boldsymbol{V}_{PR+}$. If additional positive and negative pulse responses follow, their scanning values are superimposed on one another. Therefore the values 0, +1, and -1 are formed. The 0 means that the bit has changed as compared to the previous one. By means of a known

pre-coding, it can be achieved that the result is achieved by means of dual-path rectification that -1 in +1 can be used again for a binary decision about the threshold 0.5 to 0 or 1. However, in this connection, ***3 dB*** of interference interval are lost. However, this is offset by the advantage that $\boldsymbol{H_c}(\omega)$ does not have and does not have to have a perpendicular flank like the ideal low-pass. The loss of 3 dB can be avoided by means of a Viterbi decoding.

From Fig. 7, which reproduces a partial response (duobinary code), it can furthermore be derived that one can also subtract the pulse responses of the ideal low-pass that are offset by $1/2f_g$. The related pulse response $\boldsymbol{V_{PR}}$ then has the scanning values -½ and +½. The transmission function

$$H_S(\omega) = j\sin\frac{\pi}{2}\frac{\omega}{\omega_g} \qquad -\omega \le \omega \le \omega_g$$

belongs to the subtraction of the pulse responses.

The evaluation can be carried out, as in the case of the normal duobinary signal, by means of transmission-side pre-coding and reception-side dual-path rectification. In this method step, the

bit inversion is supposed to be eliminated in the pre-coding, so that no negated bit sequence can occur. This modified duobinary coding is important for the following. In the case of ***Q^2PSK***, transmission takes place at half the bit rate f_g per channel. Accordingly, the partial response filters $\boldsymbol{H_{PR}}$ (Fig. 7) must be designed for $\omega_g/2$ instead of for ω_g, i.e. $\boldsymbol{T} = 1/2f_g$, as indicated in Fig. 8, top right. In Fig. 8, a ***Q^2PSK*** transmission is shown with partial response. In the block schematic at the top, it is indicated how the partial response filter must be subsequently switched on the reception side, in each instance. For a matched filter arrangement, it must be divided up between the transmission and reception side as $\sqrt{\boldsymbol{H_{PR}}}$. However, this can only be done for the amount in the case of $\boldsymbol{H_s}$ (Fig. 8, top right). In Fig. 8, right, the transmission functions $\boldsymbol{H_c}$ and $\boldsymbol{H_s}$ are shown in the diagram. $\boldsymbol{H_c}$ is unsuitable for ***Q^2PSK***, because this would have the result of a sign change at $f_g/2$, and an inverting filter would have to be used on the reception side from [see original for formula, page 10, line 30, second formula in that line]. In contrast, it is much simpler and easier to implement

$$\sqrt{|H_S(\omega)|} = \sqrt{\sin \pi \frac{|\omega|}{\omega_g}}$$

This function is inserted both on the transmission side and the reception side. In addition (e.g. on the reception side), a Hilbert filter with the transition function

$H_H(\omega) = j\ sign(\omega)$ can be provided, in order to be able to derive an imaginary transmission function from a real one, and vice versa. In the case of implementation of the filters by means of modulation, a cosine carrier becomes a sine carrier, and vice versa; this is shown in Fig. 9 for the example in Fig. 3a, because this example is very easy to see in an overview.

In Fig. 9, which reproduces a partial response system, the individual filters are shown at the top in a combination for a response system. In the case that $\boldsymbol{P}_1$ and $\boldsymbol{P}_2$ form a Hilbert pair, the places $\boldsymbol{P}_1$* and $\boldsymbol{P}_2$* on the reception side are simply interchanged, if the Hilbert filter is combined from $\boldsymbol{P}_1$* and $\boldsymbol{P}_2$*. A loss in interference interval is not connected with this, since the noise output and the amount of the scanning value of the wanted signal remain unchanged.

In the following, implementation by means of modulation and demodulation and transition to $\boldsymbol{Q^n PSK}$ will be described.

While $\boldsymbol{P}_1$ is a low-pass, $\boldsymbol{P}_2$, on the other hand, is a band pass. The pulse responses belonging to $\boldsymbol{P}_2\boldsymbol{P}_2{}^*$ are at a much "higher frequency" than the ones that belong to $\boldsymbol{P}_1{}^2$, as is evident from the examples in Fig. 3b - 3c. A bit rate of f_g can be transmitted in the band pass $\boldsymbol{P}_2\boldsymbol{P}_2{}^*$. In the case of the implementation of the band pass $\boldsymbol{P}_2$ by means of modulation, the carrier is not allowed to be placed in the band center of $\boldsymbol{P}_2$, which would correspond to two-sided band modulation, but instead, one must work with remaining side band modulation. This is the decisive difference as compared with Variant B, in which one would use two-sided band modulation.

Fig. 10 shows the implementation of $\boldsymbol{P}_2$ by means of modulation and reception-side demodulation of $\boldsymbol{P}_2$ and transition to $\boldsymbol{Q^nPSK}$ in the upper part, whereby a lower side band between $\omega_g/2$ and ω_g is generated by means of frequency conversion from $\boldsymbol{P}_1$ to $\boldsymbol{P}_2$. In order to have a Nyquist flank at ω_g, filtering takes place with a root Nyquist filter at ω_g, and one obtains $\boldsymbol{P}_2$. The Nyquist flank at ω_g can be different from the one at $\omega_g/2$, in principle.

In the center part of Fig. 10 (reception input) it is shown how $\boldsymbol{P}_2$ is demodulated in the low-pass range. The signal is first

sent by way of the same root Nyquist low-pass as during modulation. In this way, a Nyquist flank is achieved at ω_g. By means of demodulation and low-pass filtering with $\boldsymbol{P}_1$, one obtains the desired transmission function $\boldsymbol{P}_1^2$, by way of which transmission can take place at the bit rate f_g. It is essential that the demodulation carrier sits on the root Nyquist flank. The filter could also be a high-pass or a band pass. A band pass can easily be implemented at various frequencies by means of poly-phase filters. After the demodulation, a simple filter for suppressing the portions of the double carrier frequency is then all that is required. This method of demodulation, with poly-phase filters, is advantageous in the case of multi-carrier systems.

The root Nyquist filter on the reception side serves to generate Nyquist flanks at ω=0 after the frequency conversion, which flanks add up to a constant value in the surroundings of ω=0. However, this filter can also be combined with the reception filter, as is shown in the lower part of Fig. 10, "Combining the filters." Without the filtering on the reception side, the root Nyquist flanks would add up to a hump with the maximal value $\sqrt{2}$ at ω=0. By means of a reception filter $\boldsymbol{P}_{1E}$, however, which is

inverse in the region of the root Nyquist flanks $\omega=0$, this can be equalized.

In the case of a multi-carrier system, real and imaginary channels alternate in the case of Variant A. For the modulation and demodulation of the real channels, it is practical to perform the conversion with $\cos\omega_g t$.
In this connection, conversion should take place by way of an intermediate frequency, in order to be able to make the root Nyquist filter $\sqrt{H_R}$ the same for every channel. The conversion of the channels can, of course, take place into the high-frequency range, right away, without having to undertake another conversion with two orthogonal carriers. In this case, the orthogonal carriers are therefore completely eliminated (multi-carrier system).

The principle of dividing the transmission channel H_i up into two frequency ranges can furthermore be expanded to several frequency ranges. Fig. 11 shows this schematically for three channels, whereby the Nyquist flanks at the separation points have been left out. The center filter P_{2m} is divided up into P_2. Then, there is no cross-talk either between the channels 1 and 2 or

between the channels 2 and 3. Furthermore, there is no cross-talk between channels 1 and 3, since these are separated in terms of frequency, as long as their Nyquist flanks do not fall into one another. This method can be expanded to n channels, thereby obtaining an arrangement in the basic band, which must then be applied to a cosine carrier and a sine carrier.

Fig. 12 illustrates the transition to Q^nPSK and the overall spectrum $H_g(\omega)$, whereby the purely imaginary spectra are shown with broken lines. In this connection, the partial signals are modulated in binary manner with 0 and 1. The data in the cosine branch are designated as $d_{c1}, .d_{c2}$ ..., those in the sine branch as $d_{s1}, .d_{s2}$ ....

Since the carrier frequencies are equidistant, it is obvious to perform the modulation with IDFT and the demodulation with DFT.

As compared with an OFDM, the following advantages are obtained:

- Adjacent channels are not allowed to overlap, without there being cross-talk, because in the case of Variant A, a purely real spectrum always overlaps with a purely imaginary one having symmetrical flanks.

- In the case of OFDM, the carriers must lie very precisely in the zero passages of the si spectra that are formed by the rectangular pulse scanning, which is not critical here.

- The spectrum is compact and does not have the si runners.

- The crest factor is less, because what is being transmitted is not scanned sine and cosine vibrations, but rather pulses that die out.

- The power density spectrum of the transmission signal is constant, since the power density spectra of adjacent channels supplement one another to form a constant value, because of the Nyquist flanks.

Fundamentally, in the case of Variant B, after the demodulation and filtering, cross-talk occurs also in the basic band (as is evident from Figure 13, which illustrates the in-channel square cross-talk (IKQÜS), above) not only in the upper and lower adjacent channels, but also in the channel transmitted in squared form, in the same frequency band, in other words five times cross-talk. In the case of a distortion-free transmission

channel, however, this cross-talk is completely compensated, with an equalizer having to be used, if necessary. This cross-talk will be called in-channel square cross-talk (IKQÜS).

Also in the case of Variant A (as is evident from Figure 13, bottom), IKQÜS occurs from the overlapping on the remaining side band flank (RSB flank), in other words two times cross-talk, which compensates only in the case of distortion-free transmission. However, it is much less than in the case of Variant B, and therefore Variant A is more advantageous in this regard for certain applications (DAB, DVB-T, mobile wireless), in the case of frequency-selective channels.

In Figure 10 and also in Figure 12, the RSB flank at the carrier frequency ω_g and the other flank at the lower frequency $\omega_g/2$ have been made equal. However, it is also possible, in advantageous manner, to make the RSB flank very much steeper, as is evident from Figure 14, top (roll-off r_T after filtering with $\sqrt{H_R(\omega)}$, demodulation with ω_0, and root Nyquist flank filtering). One-sided band modulation results for $r_T = 0$. In this way, the IKQÜS can be made as small as desired, in principle. The root Nyquist flank at ω_n results in the Nyquist flank in the basic band with

roll-off r after demodulation and corresponding root Nyquist filtering. In order for the two flanks at ω_0 and ω_n not to overlap, $r + r_T$ must be ≤ 1. Figure 14 shows how one can advantageously arrange the channels in the case of multi-channel transmission. It indicates the reduction of the IKQÜS in the case of Variant A. The transmission spectrum $S(\omega)$ is indicated.

In Fig. 15, the duobinary multi-channel transmission is shown, using the example of a transmission spectrum $S(\omega)$ for the two Variants A and B. Variant B is known as a multi-carrier system from IEEE Transactions on Communications COM-15, No. 6 (December 1967) pages 805-811 as well as COM-29, No. 7 (July 1981), pages 982-989, with another solution approach. According to the known circuit, 2L basic band data with equidistant sine and cosine carriers are directly converted to the HF range by way of filters on the transmission and reception side, in two-sided band modulation with the frequencies $f_k (k=1...L)$. The spectra overlap with the Nyquist flanks, i.e. the root Nyquist flanks. On the reception side, demodulation takes place with the same carriers.

In this connection, cross-talk pulses are also obtained, but they are two different ones, namely the square ÜS (cross-talk), which

has zero places in the scanning time points, according to the invention, and the in-phase ÜS (cross-talk), which has a symmetrical spectrum around the Nyquist flank. The related time signal is multiplied by the cosine and has shifted its zero places by half the bit duration.

Implementation of the transmission-side RSB filter for the upper and lower RSB (flank below and above the carrier, respectively) advantageously takes place in the basic band, with subsequent modulation. This is shown in the following, for the root Nyquist flank with roll-off r_T.

According to Figure 16, which indicates the implementation of the transmitter-side RSB filtering, the RSB filter which is shifted into the basic band in this connection is broken down into the even and odd portion $H_g(j\omega)$ and $H_u(j\omega)$. The odd portion is multiplied by j (again, a real time function belongs to $jH_u(j\omega)$). Afterwards, conversion with a cosine carrier and a sine carrier takes place. The two portions are added or subtracted and, according to Figure 16, result in RSB filters with the RSB flank at the lower and upper band end, respectively.

If the other flank is also designed as a root Nyquist flank (roll-off r) as in Figure 15, the reception-side filtering with P_1 according to Figure 10 can be eliminated, and replaced by a simple low-pass suppression of the higher frequency portions that occur during demodulation. The two low-pass filters $H_g(j\omega)$ and $jH_u(j\omega)$ can be implemented as FIR filters, in accordance with their pulse responses.

The transmission functions that have been implemented are real. For imaginary transmission functions, as they occur in every other channel in Variant A, the cosine carrier and the sine carrier must be interchanged. This is shown in the lower part of Figure 16.

On the reception side, real and imaginary spectra overlap, as was already indicated. Figure 10 must be filtered with a root Nyquist filter. Suppression by means of compensation is not possible.

According to Figure 15, one-sided band modulation has been selected for Variant A, to generate the cos channel, which would

require a filter having a perpendicular flank. The duobinary multi-carrier transmission is shown using the transmission spectrum $S(\omega)$. In the following, it will be shown that this is not necessary.

According to Figure 17, which indicates an addition of flanks $H(\omega)$ (specifically 17a) addition to 1, 17b) addition to $H_c(\omega)$), mirror-image Nyquist flanks add up to the value 1. This can be easily shown. The Nyquist flank can be written as

$$H(\omega) = 0{,}5 + U(\omega)\ .$$

$U(\omega)$ is an odd function having the properties $U(0)=0$ and $U(\omega_\chi)=0.5$. For example, in the case of cos roll-off,

$$U(\omega) = 0{,}5 \sin \frac{\pi}{2} \frac{\omega}{\omega_\chi} \qquad -\omega_\chi \le \omega \le \omega_\chi\ .$$

Therefore

$$H(\omega) + H(\omega) = 1$$

where $\omega_\chi = r_T \cdot \omega_g$.

If the flanks are supposed to add up not to 1, but rather, as in Figure 7, on the right, to H_c, no Nyquist flanks are allowed to be used in the case of the carrier frequencies. It must be true that

$$H(\omega)+H(-\omega)=H_c(\omega)=\cos\frac{\pi}{2}\frac{\omega}{\omega_g}$$

The flank $H(\omega)$ can be determined using the approach

$$H(\omega)=\chi_1\cos\frac{\pi}{2}\frac{\omega}{\omega_g}+\chi_2 U(\omega)$$

Equation (5), inserted in Equation (4), yields $\chi_1=0.5$. If one requires, in practical manner, that $H(-\omega_\chi)=0$ and

$$H(\omega_\chi)=H_c(\omega_\chi)=\cos\frac{\pi}{2}\frac{\omega_\chi}{\omega_g},$$

therefore it follows that

$$\omega_2=\cos\frac{\pi}{2}\frac{\omega_\chi}{\omega_g}.$$

For the cos roll-off, this means that

$$H(\omega)=0.5(\cos\frac{\pi}{2}\frac{\omega}{\omega_g}+\cos(\frac{\pi}{2}\frac{\omega_\chi}{\omega_g})\cdot\sin\frac{\pi}{2}\frac{\omega}{\omega_\chi}) \qquad -\omega_\chi\le\omega\le\omega_\chi.$$

With this, the perpendicular flank can be avoided in the case of Variant A, and adjacent channels overlap. The greater the overlap, the greater the IKQÜS, of course. It is noteworthy that now, Variant A passes over into Variant B at $\omega_\chi=\omega_g$.

This method of procedure is not limited to the cos function but rather can also be used for other functions.